AN EIGHTEEN CHANNEL PAM TIME-DIVISION MULTIPLEX TELEGRAPH SYSTEM MODULATOR

BY

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-

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by

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Submitted in partial fulfillment
of the requirements
for the degree of
MASTER OF SCIENCE
in

ENGINEERING ELECTRONICS

United States Naval Postgraduate School
Annapolis, Maryland
1950

This work is accepted as fulfilling

the thesis requirements for the degree of

MASTER OF SCIENCE

in

ENGINEERING ELECTRONICS

from the
United States Naval Postgraduate School

PREFACE

This paper treats briefly the description, design considerations, performance characteristics, and applications of an eighteen channel PAM time-division multiplex telegraph system.

This work was done by the author at the Federal Telecommunications Laboratories, Inc., Nutley, New Jersey during the period January to March, 1950 while a student in the Electronics Engineering curriculum at the U. S. Naval Postgraduate School, Annapolis, Maryland.

The investigation and development came about as a result of a proposal by Messrs. D. D. Grieg and A. M. Levine of the Federal Telecommunications Laboratories for terminal equipment to be used in conjunction with the twenty-four voice channel PTM communications equipment which is manufactured by the Federal Telephone and Radio Corporation. One of these PTM equipments is used by the Keystone Pipeline Company between Philadelphia and a pumping station at Montello, Pennsylvania.

The author wishes to express his appreciation to the Federal Telecommunications Laboratories, Inc., to Messrs. D. D. Grieg, A. M. Levine, S. Moskowitz, and to the engineers and technicians of the R-6 division of FTL for the facilities provided and for the cooperation accorded him while working on this project.

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CHAPTER I

INTRODUCTION

1. <u>General Description</u>.

The eighteen channel PAM telegraph modulator is terminal equipment consisting of a timing generator, a marker or synchronizing pulse generator, and a channel modulator for each of the seventeen telegraph inputs.

The telegraph inputs may consist of any signal which has an "ON-OFF" characteristic such as land-line telegraph or teletypewriter service.

The output of the equipment is a complex audio signal which may be used to modulate a radio frequency carrier, or the modulator may be used as terminal equipment to modulate a voice channel of any multiplex communications equipment.

2. <u>Detailed Description</u>.

a. Timing Generator.

This part of the equipment consists of a closed loop circuit with positive feedback which oscillates at the base sampling frequency (183 cycles per second). The frequency of the oscillations is determined by the phase shift networks in the loop; these networks also serve the purpose of dividing the total period into equal time intervals.

Since there are an odd number of stages of amplification in the loop, the condition for oscillation will exist at the frequency for which 180 degrees phase shift (or an odd integral multiple thereof) occurs. That the stable

frequency of oscillation is the lowest frequency at which 180 degrees phase shift occurs will be shown later.

A phase shift of 180 degrees at the base frequency corresponds to a time delay of one-half the total period; neglecting time delay in the electron tubes, the nine sections of the phase shift network are inherently capable of dividing one-half the total period into nine equal time intervals; in order to divide the remaining half-period into nine equal time intervals, advantage is taken of the phase reversal which the voltage at each tap undergoes each half-cycle; thus the total period is divided into eighteen equal time intervals.

The phase relationships and the relative time in the timing generator are shown in Figure 1.

b. Marker Channel Pulse Generator. (Figure 13)

The synchronizing pulse is generated by a parallel resonant circuit which "rings" when shock excited by the signal from the plate of V_{13A}.

The first germanium crystal diode damps out any positive pulses, allowing the negative pulses to be selected. The other diode acts as a clipper to determine the amplitude of the pulse coupled to the grid of V_{13B}. The negative pulse is amplified and inverted in V_{13B}. The amplitude of the marker pulse at the output of the cathode follower, V_{13C}, is determined by the setting of the attenuator in the grid circuit of V_{13B}.

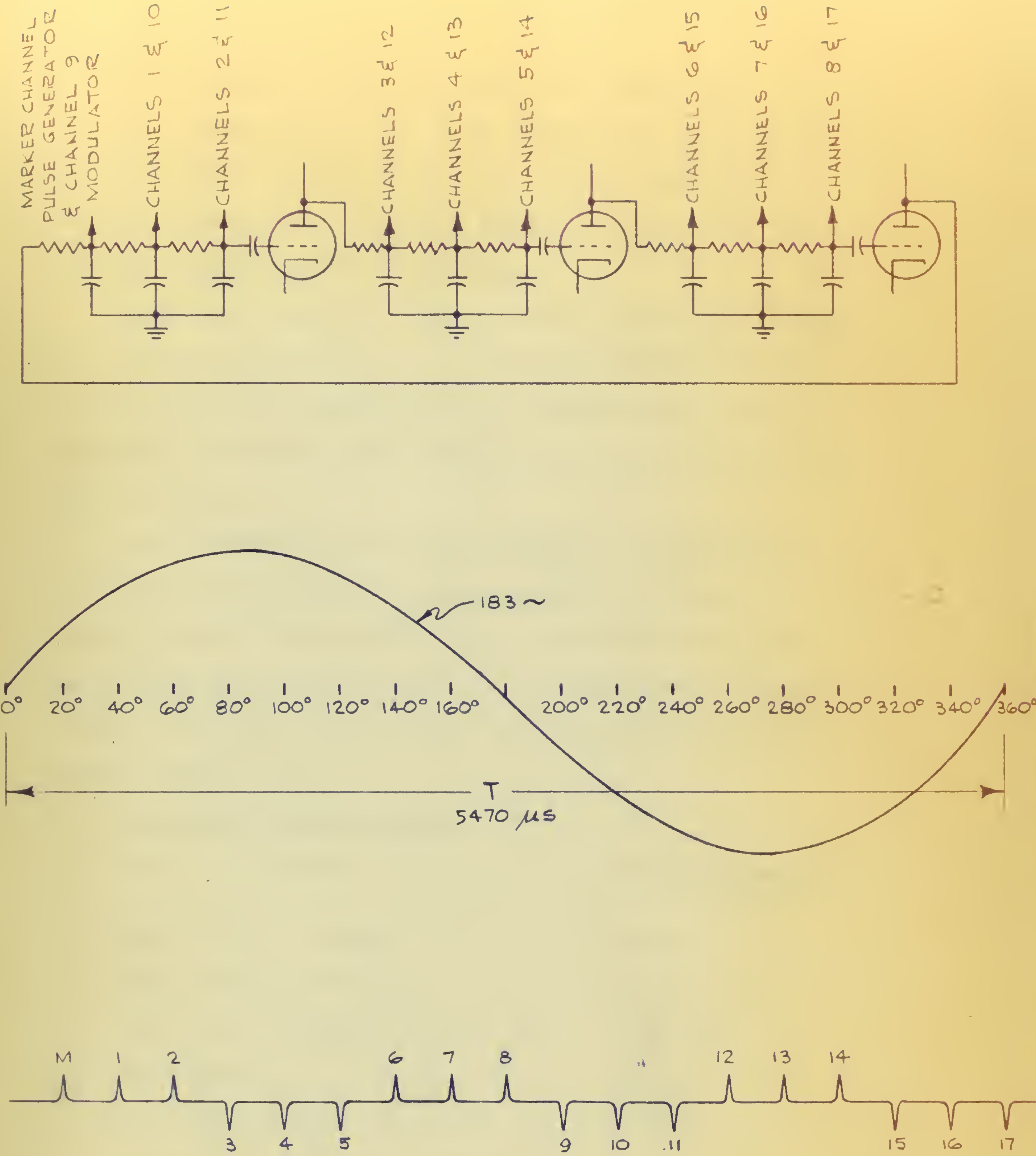


FIG. 1

RELATIVE TIME AND PHASE RELATIONSHIPS IN THE TIMING GENERATOR

c. Telegraph Channel Modulator. (Figure 15)

The pulse is formed in exactly the same manner as in the Marker Pulse Generator. In this modulator, however, the positive pulse is selected and clipped at a level of approximately three volts.

V_{15B} is a cathode follower with the cathode returned to a positive potential; the cathode follower acts as a coincidence "gate" which gives no output unless a timing pulse and a telegraph input signal are coupled to the grid simultaneously.

The schematic of the other modulator configuration for the remaining nine channels is shown in Figure 16. In this circuit the negative pulse is selected from the pulse forming network. V_{16B} is a low gain stage used as a phase inverter; the remainder of the circuit is as described above.

3. <u>Technical Specifications</u>.

Number of channels	18 (17 telegraph, 1 marker)
Bandwidth per channel	183 c/s
Base sampling frequency	183 c/s
Total pulse repetition rate	3300 c/s
Pulse width	150 μs.
Cross-talk ratio	30 db.
Telegraph signal input amplitude	10 volts.
Marker pulse output amplitude	10-15 volts.
Sampling pulse output amplitude (with 10 volt telegraph signal input)	5 volts

Output load impedance	180 ohms
Telegraph signal input impedance	.5 megohm
Power supply requirements	300 volts DC @ 300 ma. -50 volts DC @ 50 ma. 6.3 volts AC @ 10 amp.
Allowable frequency drift	± 2 %

CHAPTER II

DESIGN CONSIDERATIONS

1. <u>General Specifications</u>.

The general specifications are set forth by Federal Telecommunications Laboratories' Proposal 809, which is illustrated in Figures 2 through 7 inclusive.

Figure 2 indicates the basic function of the terminal equipment, the time-division multiplexing of eighteen telegraph signals. This specification was modified to provide only seventeen telegraph channels in order to simplify the design.

A block diagram of the terminal equipment is shown in Figure 3a, while in Figure 4 is shown a simplified schematic of the PAM telegraphy modulator.

For completeness, a simplified schematic of the demodulator is shown in Figure 5. The demodulator must consist of circuits to separate the marker pulse from the channel pulses, a delay network similar to that used in the timing generator, and a demodulator (one for each telegraph channel) which is "gated" by a pulse derived from the signal taken from the delay network. The output has the same "ON-OFF" characteristic as the telegraph input and is obtained by a correct selection of the time constant in the grid circuit of the output stage.

It is necessary that the PAM complex signal be confined within the bandwidth of one voice channel (3.3 kilocycles/sec.). Since the pulse output of the

BLOCK DIAGRAM OF OVERALL 18 CHANNEL PLUS MARKER TELEGRAPHY SYSTEM UTILIZING TIME DIVISION

USED WITH	ASSEMBLY	
PROTOCAL BOX		

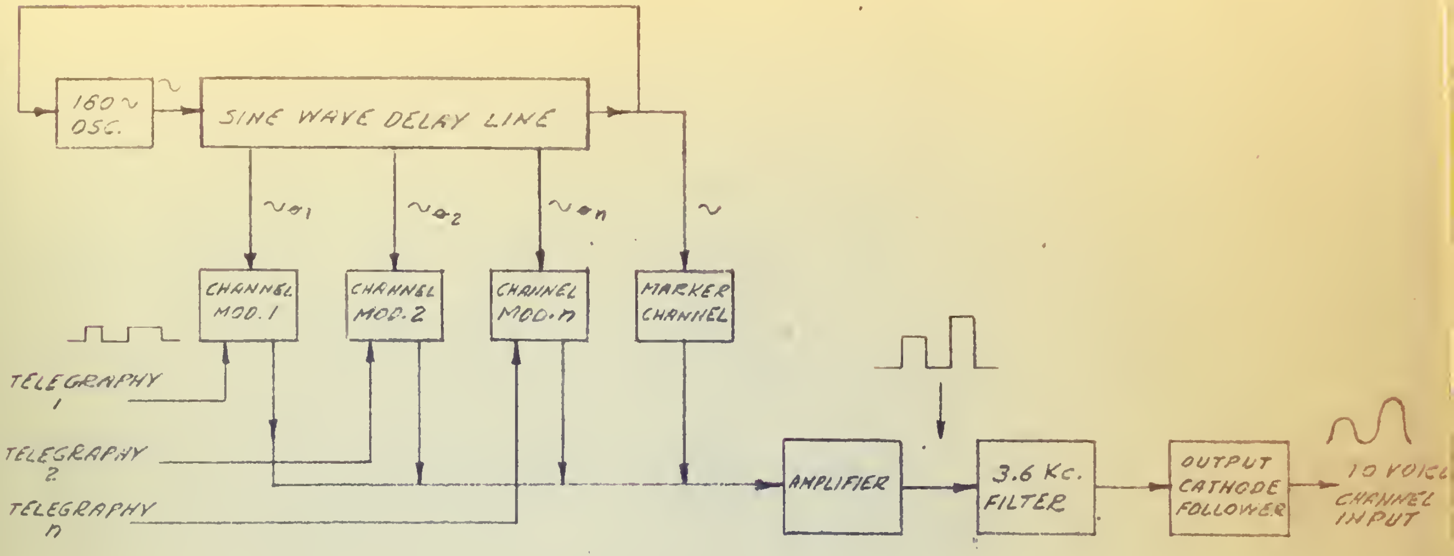
Federal Telecommunication Laboratories, Inc.
NUTLEY, N. J.

SHEET OF

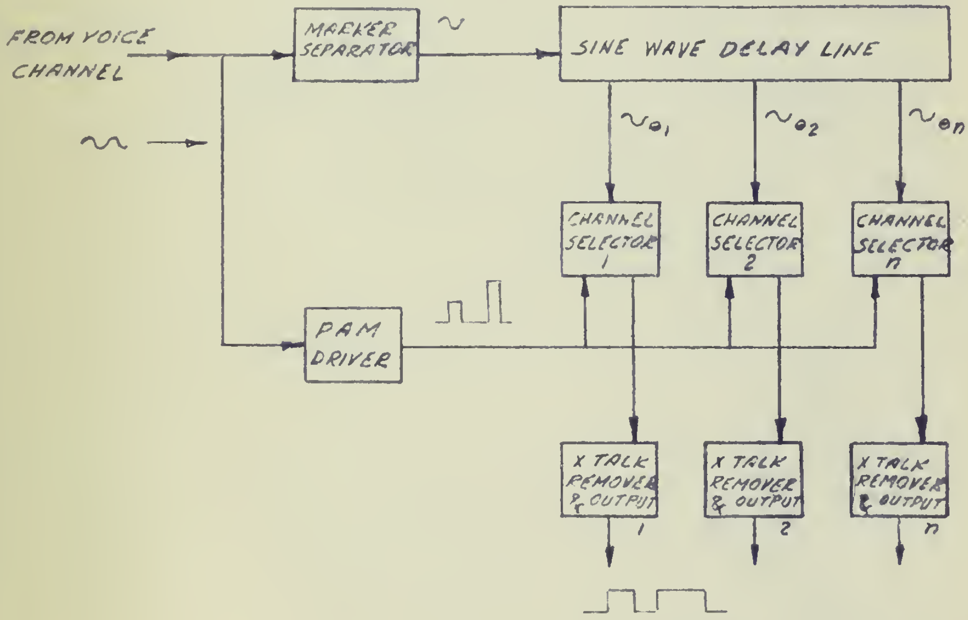
APPD	ENG	CKD	DRAWN

FIG. 2

KA-3716-[illegible]



a. BLOCK DIAGRAM OF 18 CHANNEL TELEGRAPHY MODULATOR



b. BLOCK DIAGRAM OF 18 CHANNELS TELEGRAPHY DEMODULATOR

USED WITH	ASSEMBLY
PROPOSAL 809	

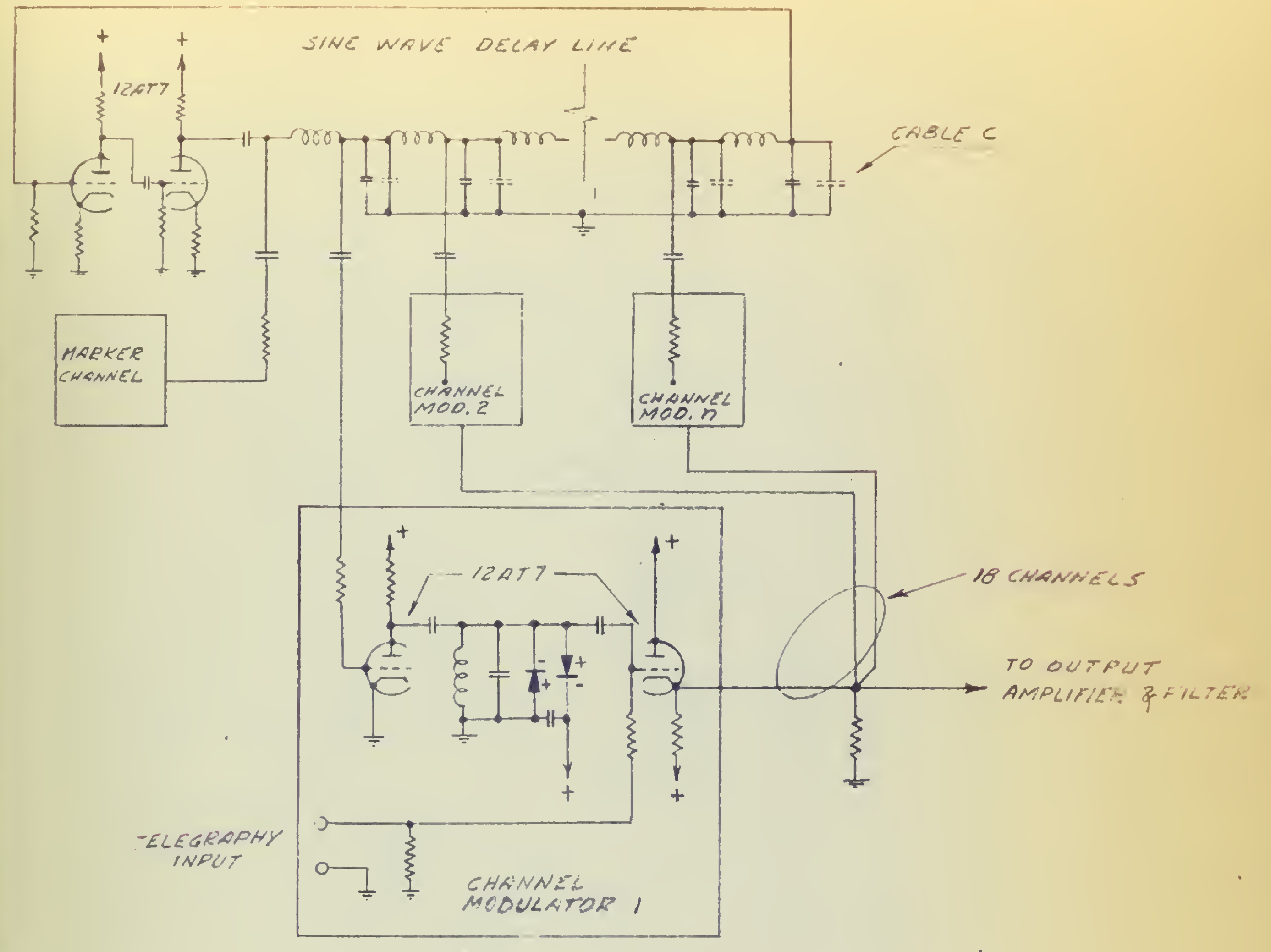
Federal Telecommunication Laboratories, Inc.
NUTLEY, N. J.

SHEET OF

APPD	ENG.	CKD	DRAWN
			JF.

FIG. 3

RX-371841-1A



SIMPLIFIED SCHEMATIC OF 18 CHANNEL
PAM TELEGRAPHY MODULATOR
(ONLY ONE TYPE OF TUBE USED: 12AT7)

USED WITH PROPOSAL 809	ASSEMBLY	THIS DRAWING IS THE PROPERTY OF FEDERAL TELECOMMUNICATION LABORATORIES INC. THE INFORMATION CONTAINED IN IT OR SUPPLIED WITH IT IS GIVEN IN STRICT CONFIDENCE AND MAY BE USED ONLY FOR FILLING ORDERS FOR FEDERAL TELECOMMUNICATION LABORATORIES INC. ON DEMAND THIS DRAWING MUST BE RETURNED TO FEDERAL TELECOMMUNICATION LABORATORIES INC. NO COPIES MAY BE MADE WITHOUT THE CONSENT OF FEDERAL TELECOMMUNICATION LABORATORIES, INC.
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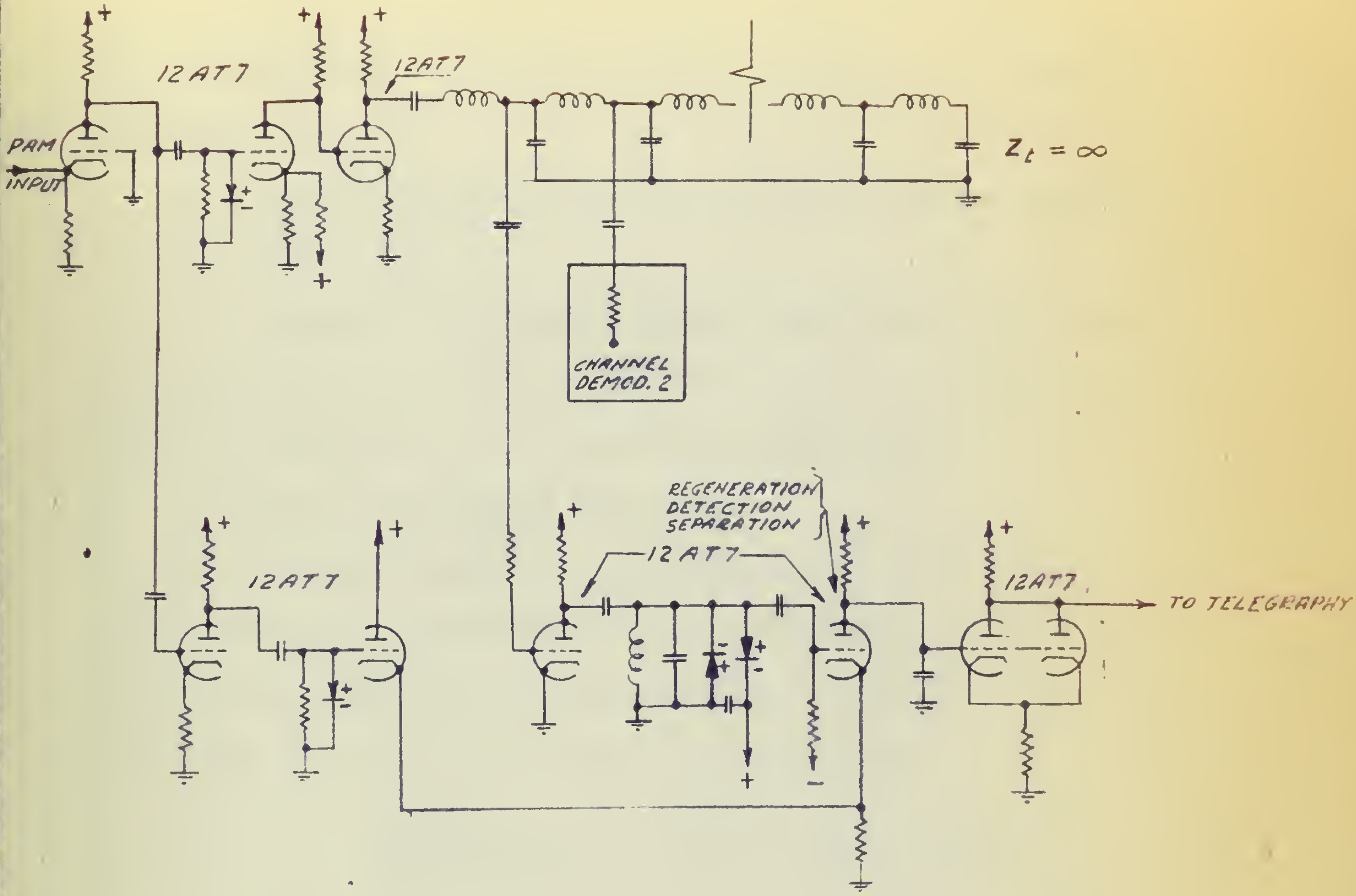
SHEET OF		FIG. 4
APPD / ENG.	CKD. / DRAWN JF	RX-371842-1A



SIMPLIFIED SCHEMATIC OF 18 CHANNEL
PAM TELEGRAPHY DEMODULATOR

ONLY ONE TYPE OF TUBE USED (12AT7)

USED WITH PROPOSAL 809	ASSEMBLY	

Federal Telecommunication Laboratories, Inc.
NUTLEY, N. J.

SHEET OF

APPD	ENG	CKD	DRAWN
			JF.

FIG. 5

RX-371843-1A

modulators and marker pulse generator have been shaped and have relatively steep sides, the signal will occupy a bandwidth greater than that available. The mixed signal must, therefore, be passed through a low pass filter before modulating the voice channel. This is illustrated in Figure 6.

The physical limitations on the equipment are specified in Figure 7.

The design of the plate loaded amplifiers and cathode followers was done graphically from the tube characteristics and is considered to be conventional procedure.

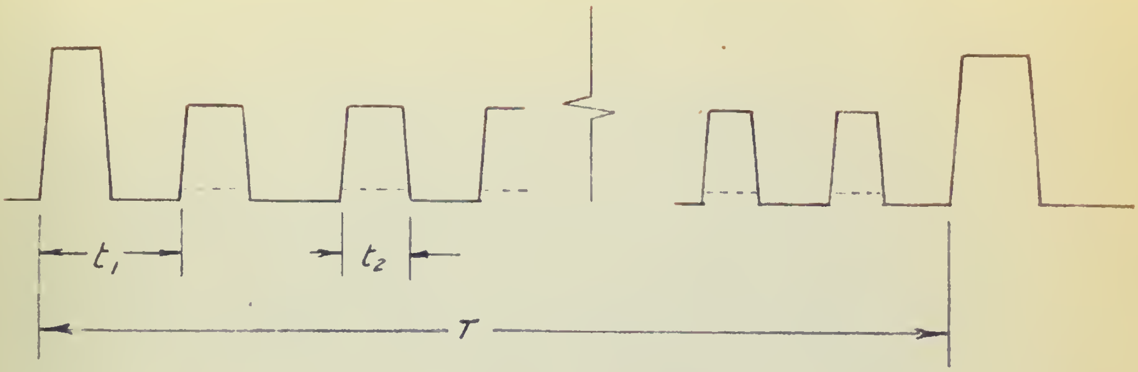
2. <u>Design of the Timing Generator</u>.

The bandwidth available in each voice channel to be modulated is approximately 3.3 kc. Eighteen channels of intelligence are required within this bandwidth. The bandwidth available per channel is therefore:

$$\frac{3300 \text{ c/s}}{18 \text{ channels}} = 183 \text{ c/s per channel}$$

The telegraph input signal has a quantized waveform, i.e., it has an "ON-OFF" characteristic, although it may consist of long or short pulses and of spaces in order to contain the intelligence. This input signal may be represented in the extreme case by a series of short pulses. For example, at the maximum telegraph speed specified (60 w.p.m.) the repetition rate is 25 p. p. s.

The sampling frequency is made as high as possible without exceeding the available bandwidth by making it equal to the channel bandwidth (183 c/s).



WAVE FORM OF PAM BEFORE GOING THRU OUTPUT FILTER
$t_1 = 29.2\ \mu sec.$, $t_2 = 14.6\ \mu sec.$ and $T = 5500\ \mu sec.$

WAVE FORM OF PAM, SHOWING THE SAME TIMING AS THE ABOVE, BUT PASSED THRU A LOW PASS FILTER.

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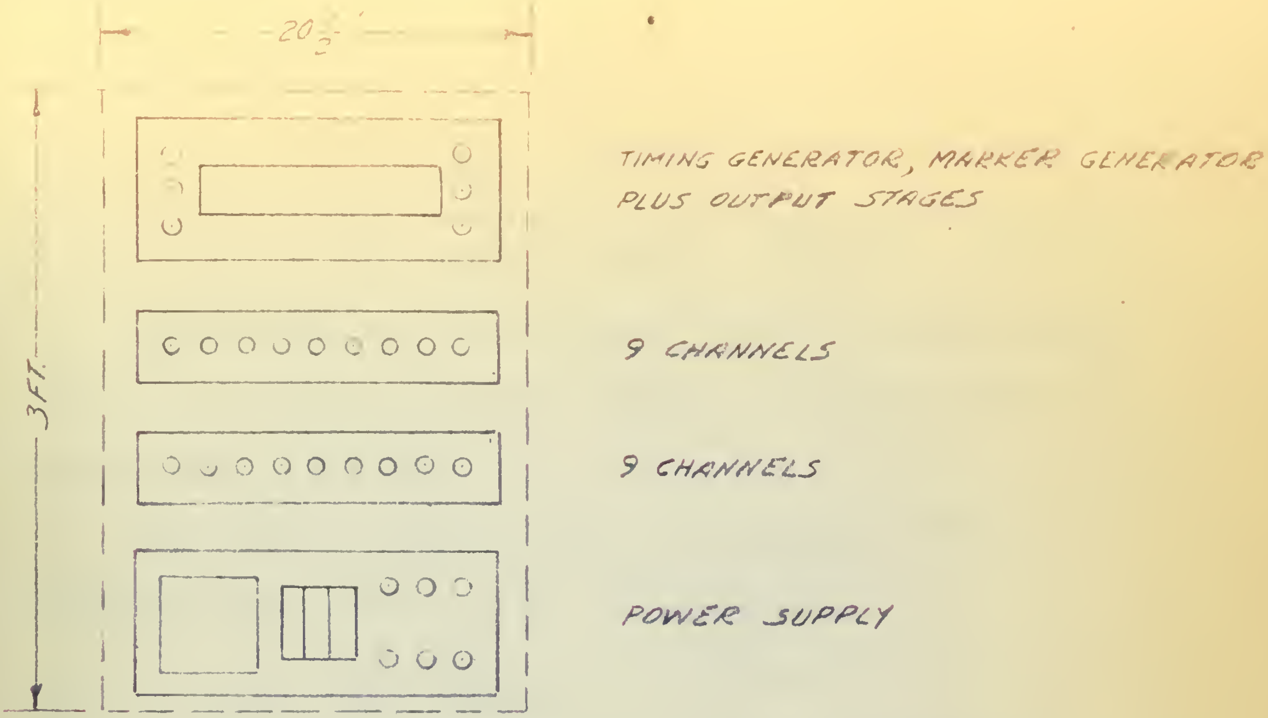
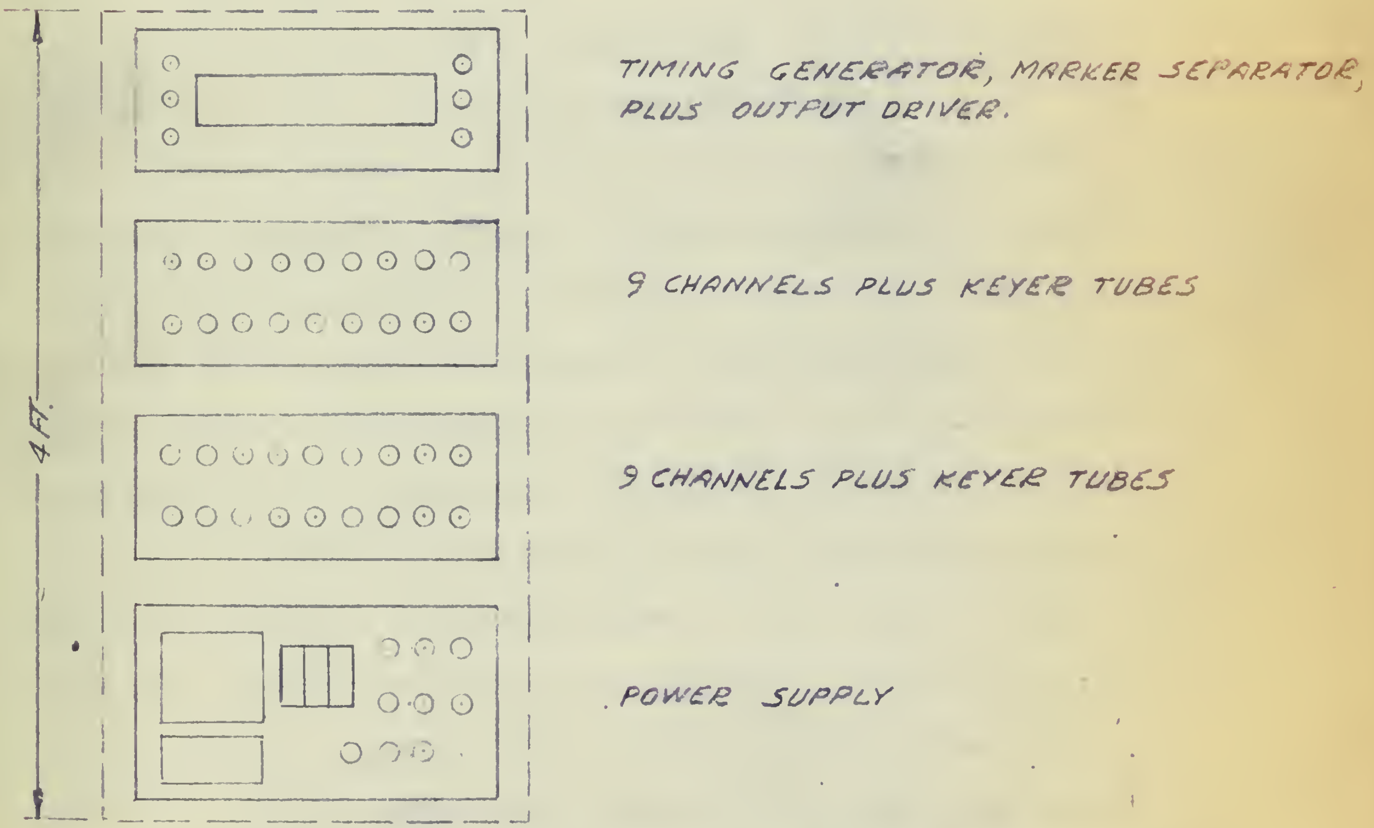
		Federal Telecommunication Laboratories, Inc. NUTLEY, N. J.
SHEET OF		FIG. 6
APPD ENG CKD. DRAWN JF.		(12) RX-371844-A



PHYSICAL LAYOUT OF MODULATOR



PHYSICAL LAYOUT OF DEMODULATOR

USED WITH	ASSEMBLY
PROPOSAL 809	

Federal Telecommunication Laboratories, Inc.
NUTLEY, N. J.

SHEET		OF	
APPD.	ENG.	CKD.	DRAWN
			JF

FIG. 7

RX-371845-1A

The total period is then:

$$T = \frac{1}{183} = 5470\,\mu s.$$

$$\tau = \text{channel width} = \frac{5470\,\mu s.}{18} = 303\,\mu s.$$

After mixing (in the extreme case), eighteen pulses will occur within the total period, uniformly spaced by approximately 303 μs.

The bandwidth required for transmission of the fundamental is then:

$$\Delta f = \frac{1}{303\,\mu s.} = 3300 \text{ c/s}$$

Since $\frac{183}{25} \cong 7.3$, the sampling rate is considerably faster than is necessary to reproduce the 60 w. p. m. telegraph signal with negligible distortion.

Theoretically the sampling rate must be twice the highest frequency component of the waveform to be sampled if it is to be reproduced without distortion. Practically, a ratio of 2.5 has been found to be satisfactory. In this application it is necessary to transmit only the fundamental in order to detect the presence or absence of a pulse. The theoretical sampling rate of twice the highest frequency component to be reproduced is necessary so that the sum and difference frequencies produced by the modulation process can be removed from the waveform by a low-pass filter. This is important in the demodulation process to prevent distortion and cross-talk. (Moskowitz, S., 11)

It was desirable for the timing generator to be small physically with a minimum number of vacuum tubes. It also was specified that any drift in frequency not cause channel timing pulses to overlap. This requirement dictated the use of a timing device which itself would determine the oscillator frequency.

In Figure 8 is shown the circuit of an early design to fulfill the above requirements. The timing device and frequency determing element was a lumped constant delay line. The delay line was in the form of a constant "k" type filter and was broken in the middle to insert a two stage amplifier.

The characteristics of this circuit are interesting and serve to illustrate certain factors leading to the final design.

Note that there are an even number of stages of amplification. An oscillatory condition exists, therefore, when a 360 degree (or multiple thereof) phase shift occurs. This circuit displays a tendency to oscillate in three modes: 1, a multivibration, the frequency of which is determined by the time constant of the coupling networks; 2, an oscillation determined by the phase shift of the coupling networks, the phase shift of the delay lines being small; and 3, an oscillation determined principally by the phase shift of the delay lines, the phase shift of the coupling networks being negligible at this frequency.

The input impedance of the delay lines was of the

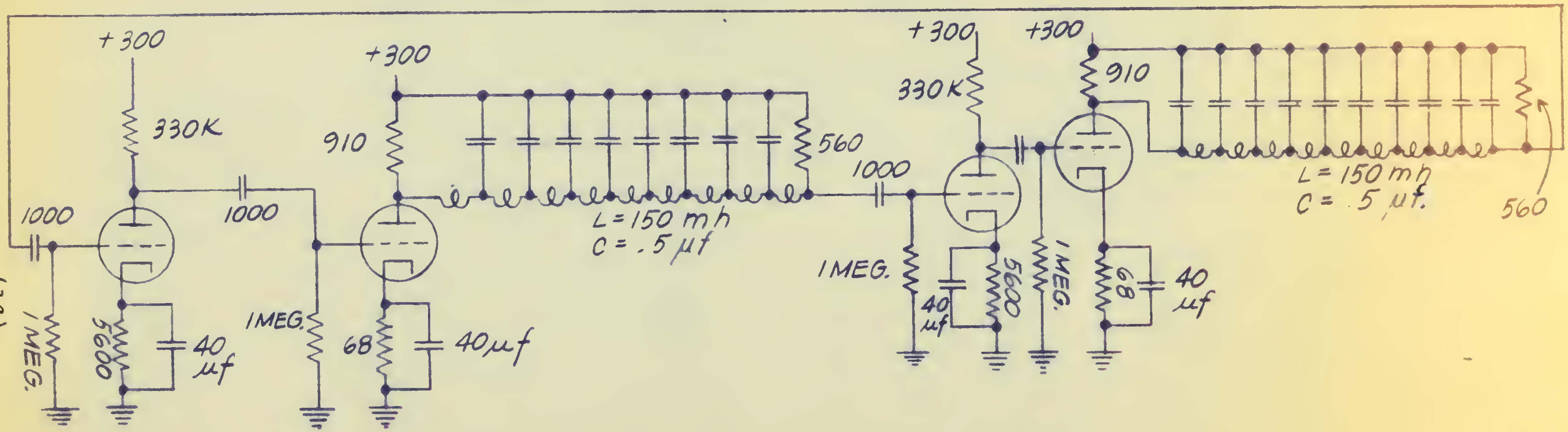


FIG. 8
EARLY DESIGN - TIMING GENERATOR

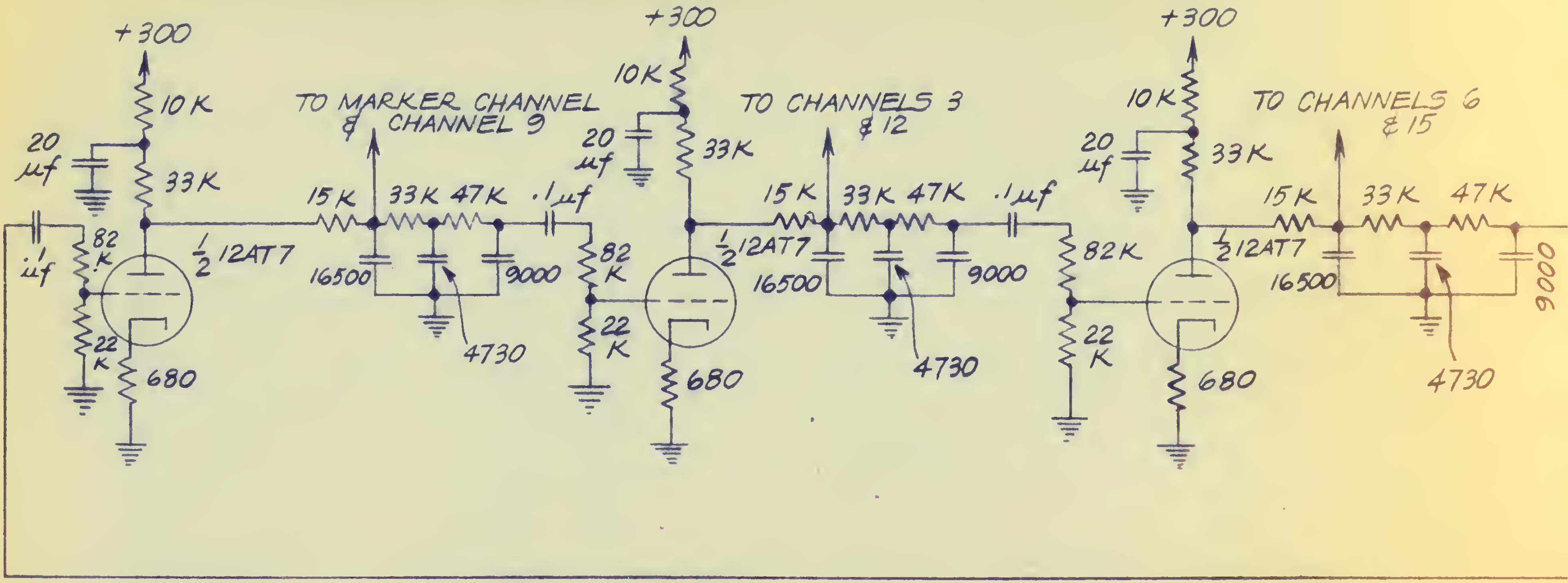
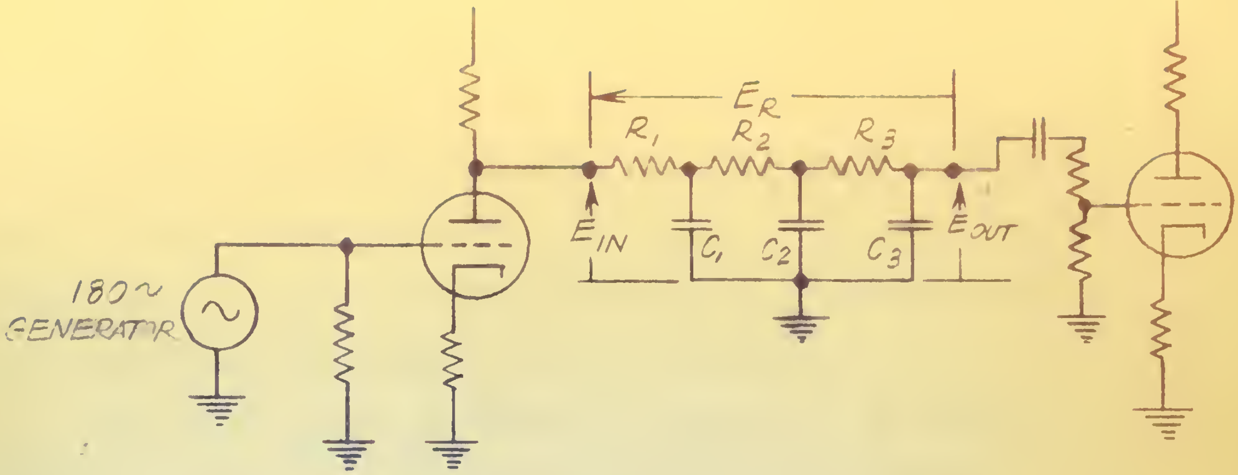
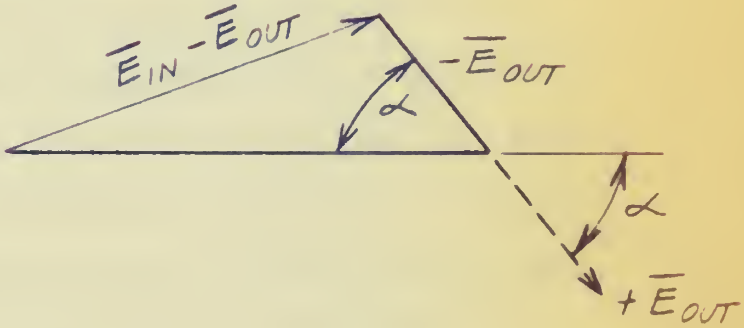


FIG. 9
183 ~ TIMING GENERATOR





$$\left(E_{IN} - E_{OUT}\right)^2 = \left(E_{IN}\right)^2 + \left(E_{OUT}\right)^2 - 2\, E_{IN} E_{OUT} \cos\alpha$$

$$\cos\alpha = \frac{\left[\left(E_{IN}\right)^2 + \left(E_{OUT}\right)^2\right] - \left[\left(E_{IN} - E_{OUT}\right)^2\right]}{2\, E_{IN} E_{OUT}}$$

FIG. 10
METHOD OF MEASURING ACCURATELY THE PHASE SHIFT OF THE NETWORKS AT VERY LOW FREQUENCIES.

order of 500 ohms when components of reasonable size were used, which limited the gain realizable in V_{8B} and V_{8D} to about two. The attenuation in the delay lines was also very high (approximately 3 db. per section).

The final design is shown in Figure 9. RC phase shift networks are used as the frequency determining elements. The shunt capacity type of network was used to make the network "low-pass", which precludes the possibility of a higher frequency mode being stable. This configuration also reduces the size of the capacitors and minimizes the effect of stray and wiring capacities.

An odd number of stages is used to prevent the very low frequency modes mentioned previously from occuring. The stable frequency of oscillation will be the lowest frequency at which the correct phase shift occurs to maintain oscillations.

The method of measuring the phase shift is indicated in Figure 10. The networks were calculated approximately as shown below and then padded to make the phase shift exact.

In a single RC section:

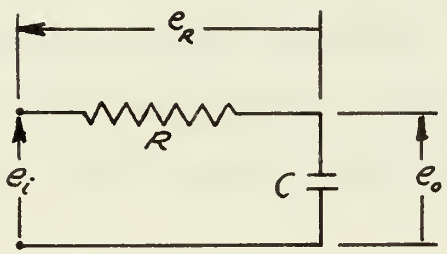


FIG. 11

$$\theta \equiv \text{phase angle}$$

$$\tan\theta = \frac{e_R}{e_o} = R\omega C$$

$$f = 183 \ \sim/s$$

$$\omega = 1150$$

$$\theta = 20^\circ$$

$$RC = \frac{\tan\theta}{\omega} = 3.17 \times 10^{-4}$$

If a suitable value of R or C is chosen, then the other may be calculated. Each succeeding section is constructed of components having larger impedances than the preceding section to minimize the loading effect.

To prevent over-driving the tubes, some attenuation was necessary at the input to each stage. The attenuators must be selected such that the grid resistance is low enough to introduce no appreciable phase shift with the grid to plate capacitance of the tube; the magnitude of the coupling resistance must, at the same time, be large enough to not appreciably load the phase shift network or introduce (with the coupling capacitor) a phase shift in the opposite direction which is not negligible.

Cathode degeneration is used because this improves the waveform. It is important to have a sinusoidal waveform since the phase shift in the networks will be as designed only to an impressed sinusoidal voltage.

De-coupling filter networks are used in the plate circuits to prevent low frequency "motor-boating".

3. Design of the Marker Pulse Generator. (Figure 13)

The desired pulse width is approximately half the channel width. This choice reduces the frequency stability requirements as well as the output filter requirements, however, there is also a decrease in the cross-talk ratio.

$$T = \text{total period} = 5470\ \mu s.$$

$$\tau = \text{channel width} = 303\ \mu s.$$

$$\frac{\tau}{2} = \text{pulse width} = 151.5\ \mu s.$$

A simple parallel resonant circuit which is shock excited into oscillation is used to generate a pulse of the desired width.

Neglecting resistance:

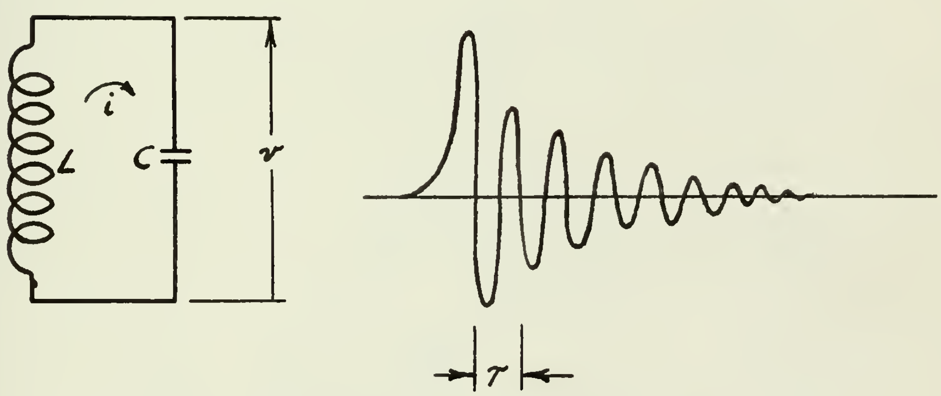


FIG. 12

$$\tau = 2\pi\sqrt{LC} = 303\ \mu s.$$

$$LC = 2.33 \times 10^{-9}$$

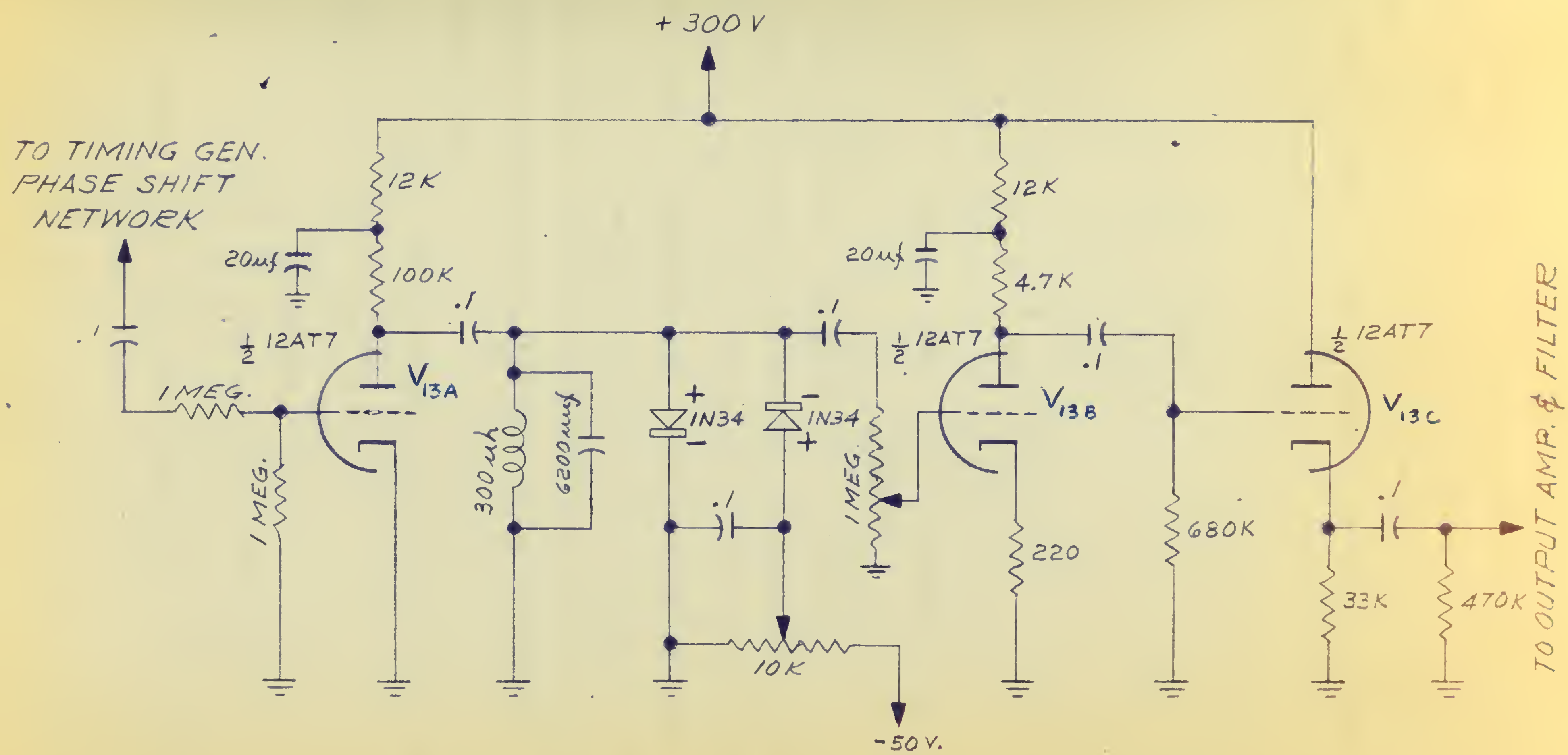


FIG. 13

MARKER CHANNEL PULSE GENERATOR

The amplitude of the pulse is a function of the L/C ratio and the slope of the sine wave signal on the grid of V_{13A}. This may be shown by assuming no dissipation during the first cycle; equating the energy stored in the electric and magnetic fields:

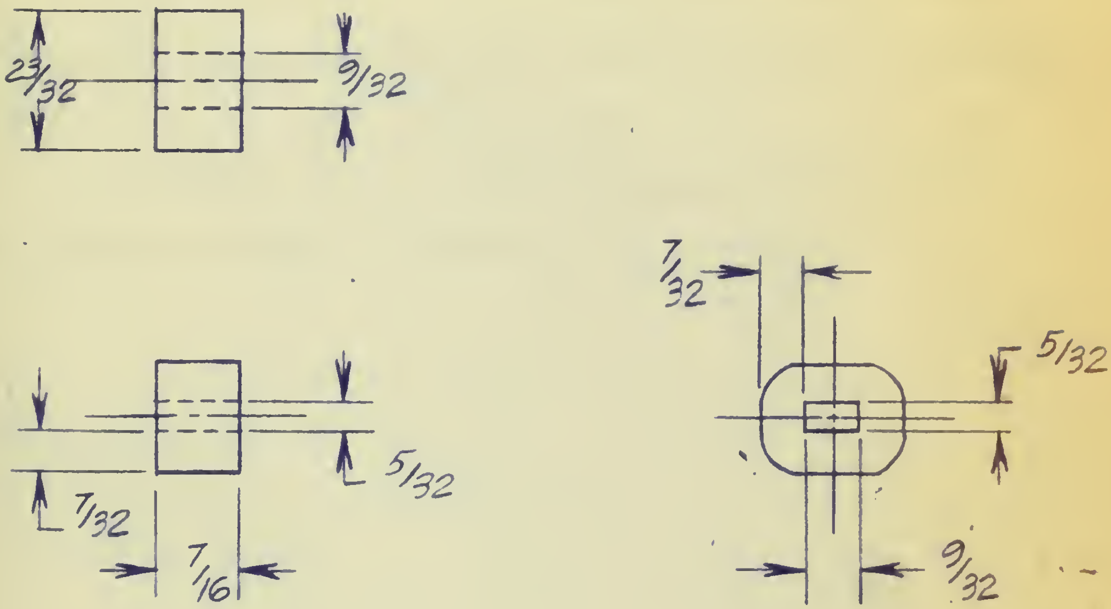
$$U_H = U_E$$
$$\tfrac{1}{2}LI^2 = \tfrac{1}{2}CV^2$$
$$V^2 = I^2 \, L/C$$
$$V = I\sqrt{\tfrac{L}{C}}$$
$$\therefore \quad V_{max} = I_{max}\sqrt{L/C}$$

But, $i = C\,\frac{dv}{dt}$, and $\frac{dv}{dt}$ is a function of the time rate of change of the signal applied to the grid of V_{13A}. It is evident that the amplitude of the sine wave input to the modulator should be as large as possible to have a steep slope when going through zero; the L/C ratio should be as large as practicable. In selecting the load resistance of the pulse generator tube, V_{13A}, it should be borne in mind that this load resistance in parallel with the plate resistance of the tube acts to damp the oscillations.

The design specifications for the inductance are indicated in Figure 14.

$$L = 300 \text{ mh.}$$
$$\therefore \quad C = 7.75 \times 10^{-9} = 7750\ \mu\mu\text{f.}$$

A value of $C = 6200\,\mu\mu f.$ was found to be satisfactory.



\# 34 FORMEX WIRE
FISH PAPER CORE
800 TURNS
DC RESISTANCE - 20 OHMS
L = 300 mh.

FIG. 14
INDUCTANCE WINDING
FOR HYPERSIL CORE

4. <u>Design of a Channel Modulator</u>.

There are two modulator configurations as shown in Figures 15 and 16. The sampling pulses are formed exactly as in the marker pulse generator and differ only in amplitude. The amplifier stages and cathode follower output stage are of conventional design. The sampling pulses and the telegraph input signal are added linearly in the grid resistance of the cathode follower coincidence "gate".

The waveforms in the pulse forming section of the modulators are shown in Figure 17.

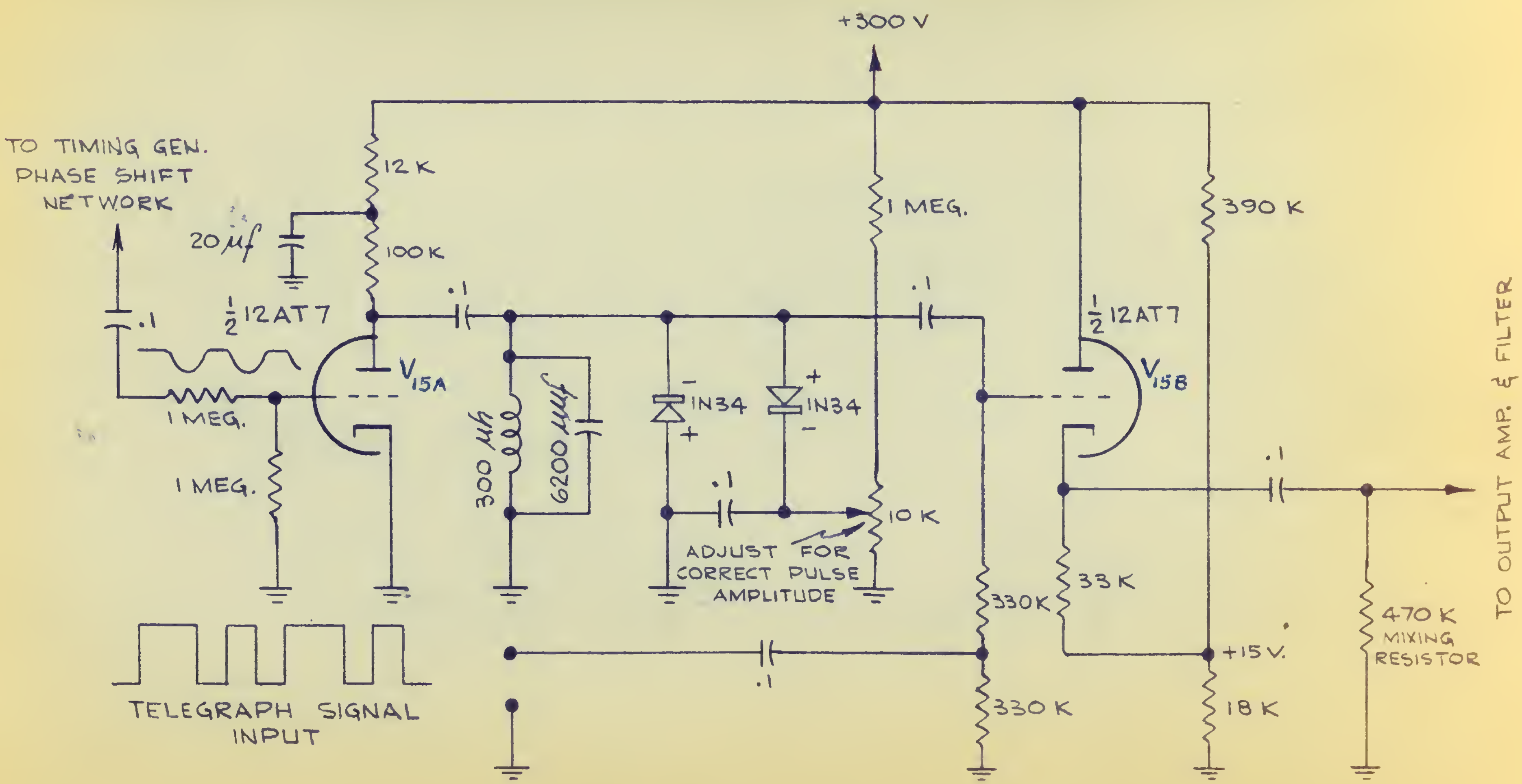


FIG. 15

MODULATOR SCHEMATIC
FOR CHANNELS 1, 2, 6, 7, 8, 12, 13, AND 14

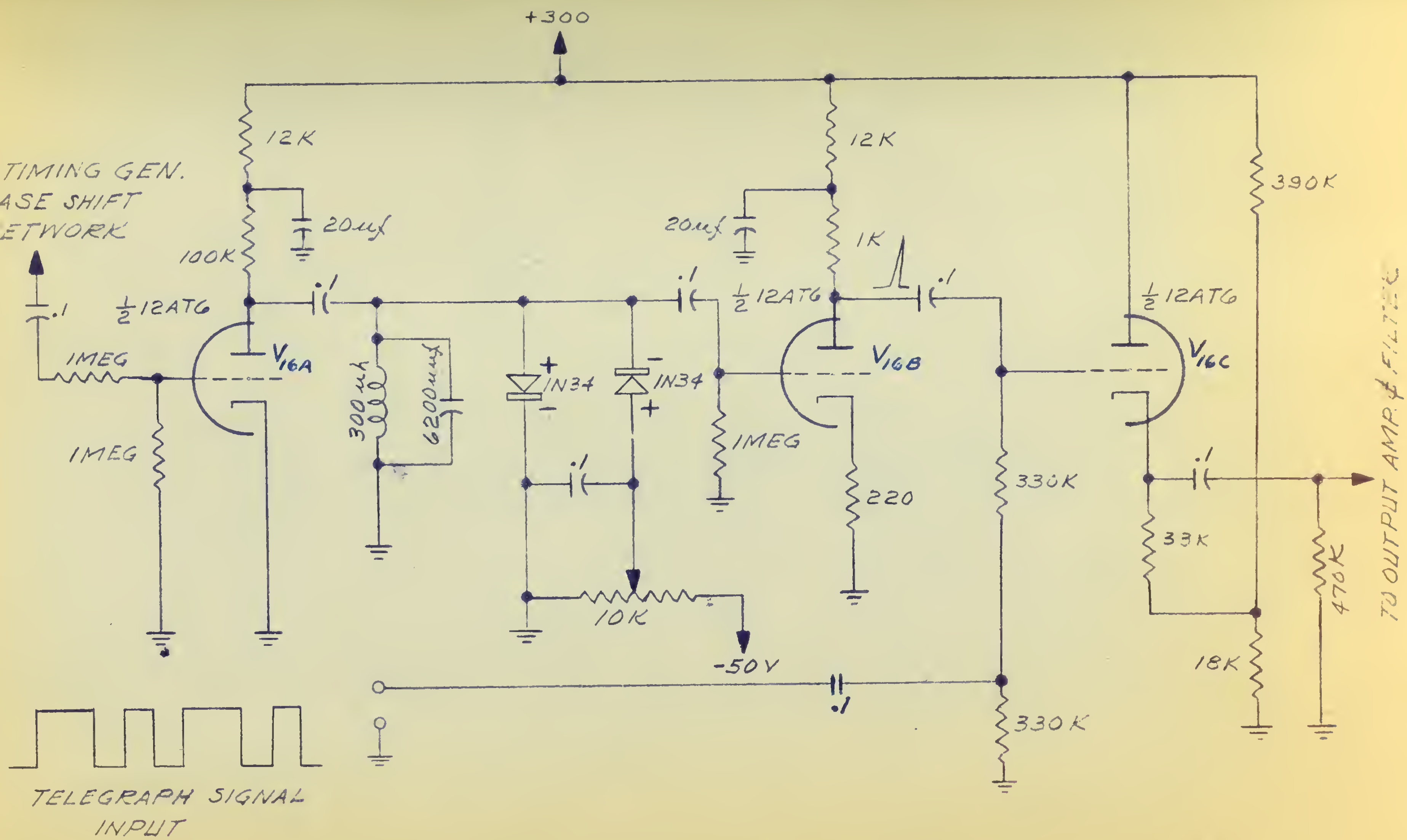


FIG. 16

MODULATOR SCHEMATIC FOR CHANNELS
3, 4, 5, 9, 10, 11, 15, 16, & 17.

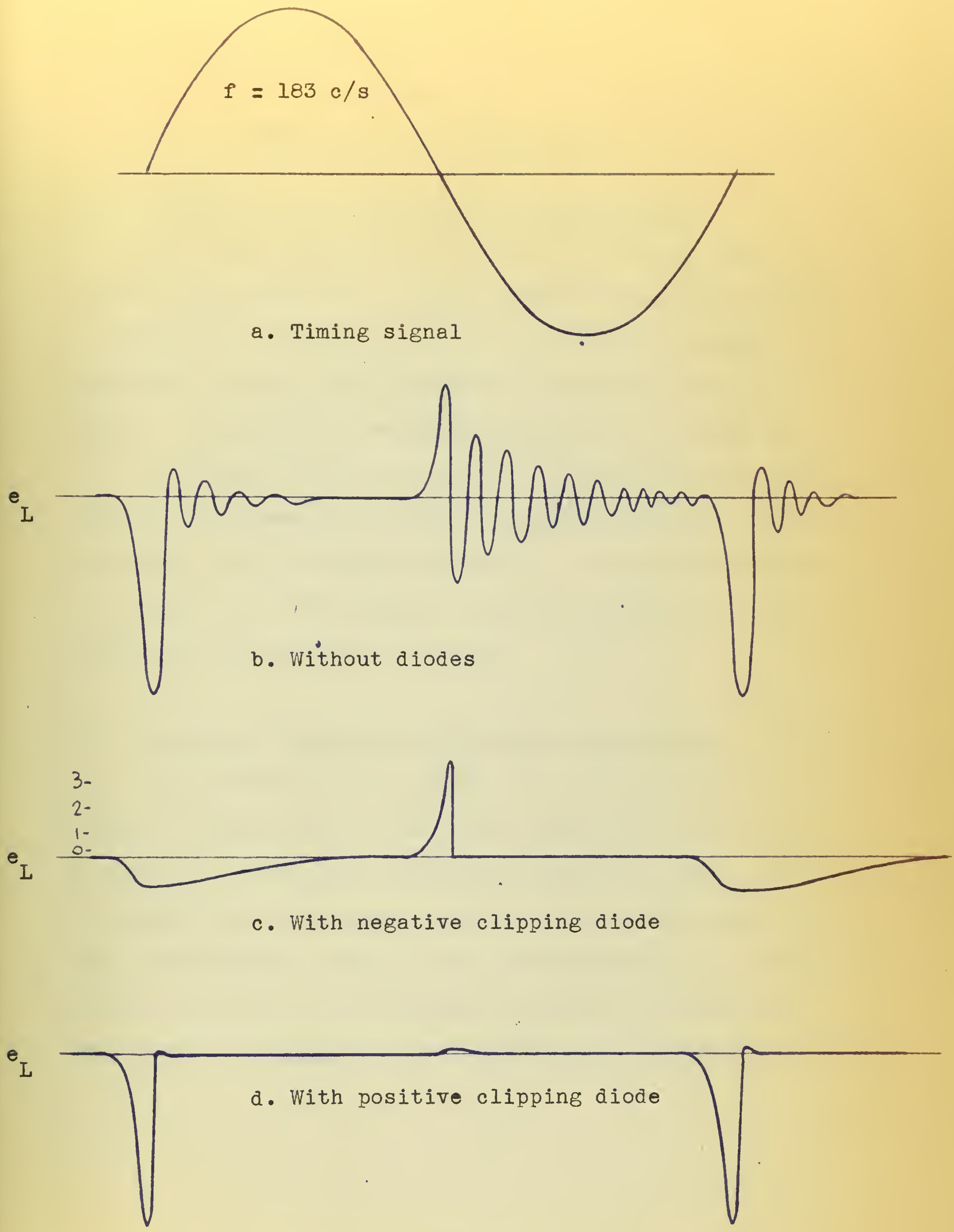


FIG. 17

WAVE FORMS IN MODULATOR RINGING CIRCUIT

CHAPTER III

PERFORMANCE CHARACTERISTICS

1. Frequency Drift With Change in Temperature.

The timing generator was placed in a Tenney test chamber and subjected to ambient temperatures from -20° C. to +50° C. A Hewlett Packard 206 A calibrated signal generator was used as a standard in measuring the frequency. The results are shown graphically in Figure 18.

No attempt was made to make the frequency drift small since to determine the maximum amount of frequency drift the system could tolerate was one of the purposes of this investigation. Five percent tolerance components were used in the phase shift networks.

2. Response to Simulated Telegraph Signal Input.

A sine wave with a frequency of approximately 30 c/s, corresponding to the pulse repetition rate of a 72 w. p. m. telegraph signal was applied to the input of channel two modulator. A sine wave was used rather than a rectangular wave in order to demonstrate the PAM characteristics of the response to inputs with varying amplitudes. The response is illustrated by Figure 19.

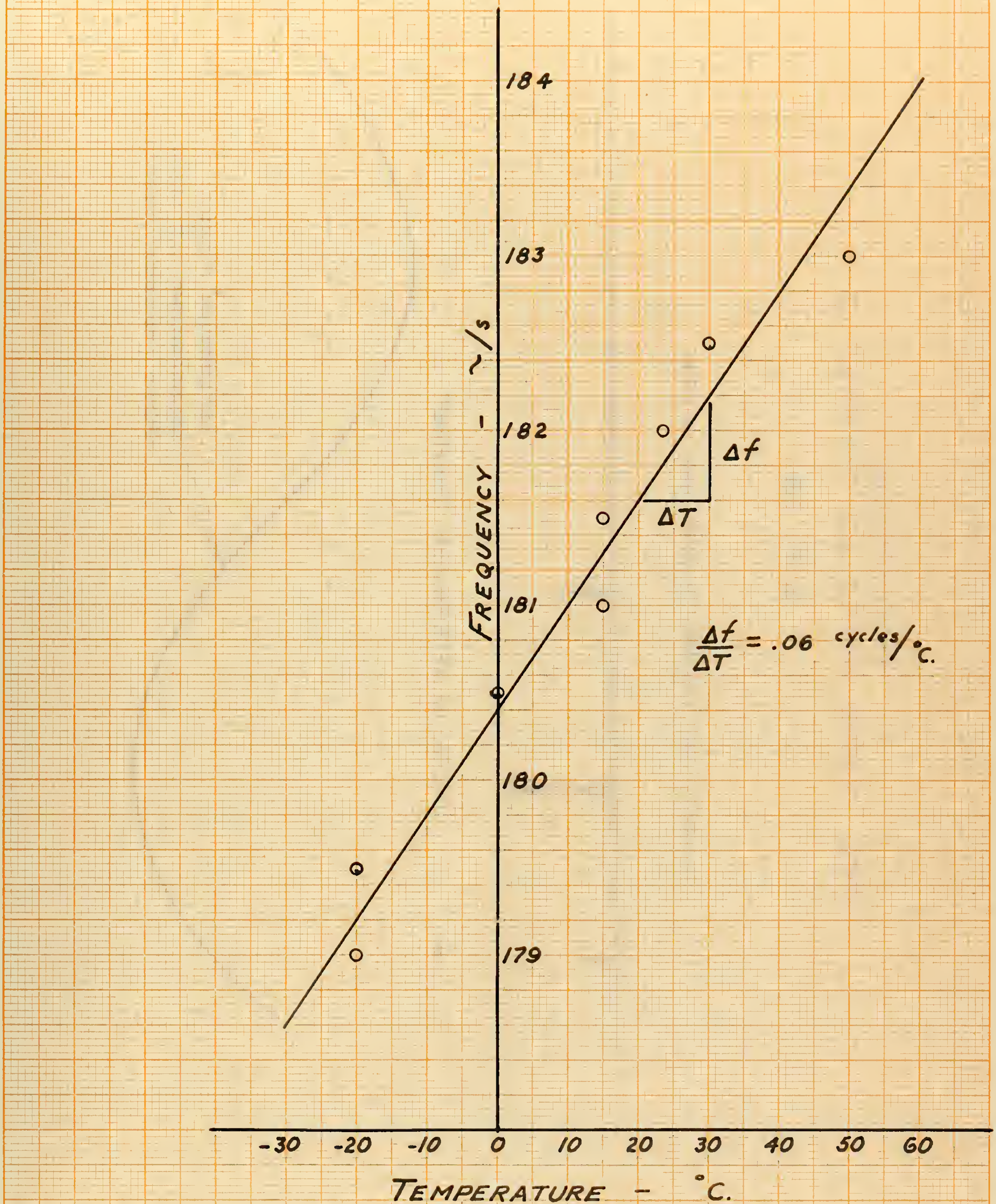

184
183
182
181
180
179
FREQUENCY - ~/s
Δf
ΔT
Δf/ΔT = .06 cycles/°C.
-30
-20
-10
0
10
20
30
40
50
60
TEMPERATURE - °C.

FIG. 18

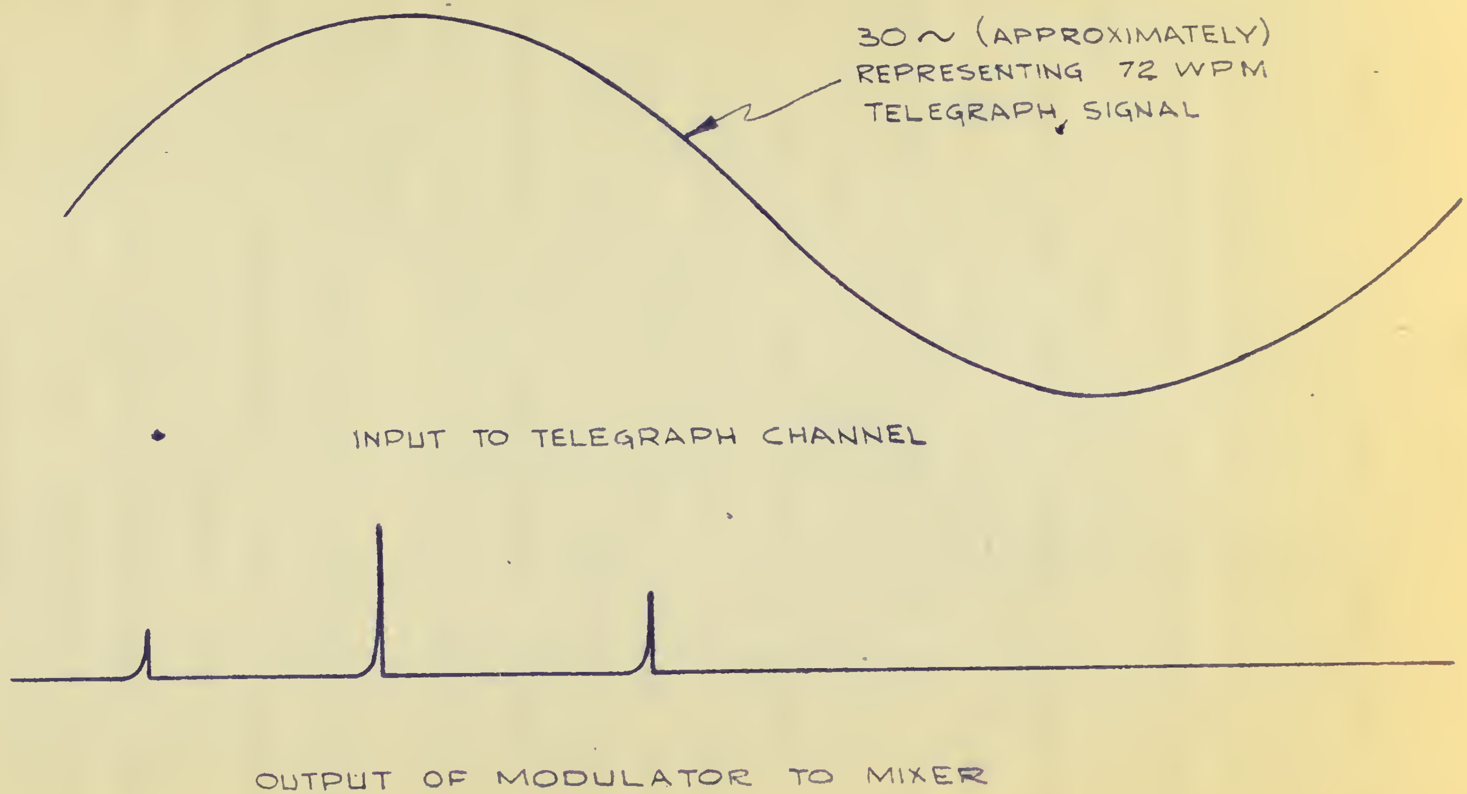


FIG. 19

SINE WAVE INPUT TO CHANNEL 2 MODULATOR

CHAPTER IV
CONCLUSIONS

In designing a low frequency phase shift oscillator suitable for the timing device and sampling rate generator in a time-division multiplex system, it is preferable to use an odd number of stages of amplification to prevent extremely low frequency multivibrator action. The phase shift network should then be designed to give 180 degrees phase shift at the desired frequency of oscillation. If the phase shift network is made to act as a low-pass filter, the stable oscillatory frequency will be the lowest at which 180 degrees phase shift occurs.

The LC lumped constant delay line is undesirable as the frequency determining element for very low frequency oscillators because of the low input impedance and the high attenuation. The impedance and transfer characteristic may be improved by sacrificing space, weight, and cost requirements.

The RC phase shift network using shunt capacity appears to be satisfactory for this application. The frequency stability may be improved by the use of precision components and some temperature compensation, however, the system is quite tolerant of frequency drift and is within the theoretical maximum allowable. (See Figure 18 and Appendix I).

Although the inference throughout has been to

telegraphic communication or similar service, the system lends itself admirably to remote control in certain industrial applications where multiplex voice communications equipment is used. For example, remote control of rotating machinery and the dam gates in hydro-electric installations.

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APPENDIX I

CALCULATION OF FREQUENCY STABILITY REQUIREMENTS

Assuming the base sampling frequency to be 183 c/s, that the timing device in the demodulation equipment gives the correct delay at this frequency, and assuming that this device has constant characteristics:

$$T_o = \text{ total period } = \frac{1}{183} = 5470\,\mu s.$$

$$\tau_o = \text{ channel width } = \frac{5470\,\mu s.}{18} = 303\,\mu s.$$

If the frequency of the base oscillator drifts, the channel width will not remain constant but will remain uniform; the effect will be to shift each channel pulse in time a progressive amount, using the marker pulse as a reference. The last channel pulse will be shifted the greatest amount in either direction, the direction depending upon the direction of the frequency drift.

Stipulate that the last channel may shift 80 % of half of a channel width and still be received by the demodulator in the correct channel:

$$\Delta T = \frac{303\,\mu s.}{2} \times (.80) = \pm 121\,\mu s.$$

$$T' = T_o + \Delta T = 5470 + 121 = 5591\,\mu s.$$

$$f' = \frac{1}{T'} = 179\,\mu s.$$

$$\Delta f = -4 \text{ c/s}$$

$$T'' = T_o + \Delta T = 5470 - 121 = 5349\,\mu s.$$

$$f'' = \frac{1}{T''} = 187\,\mu s.$$

$$f = +4 \text{ c/s}$$

Allowable drift $= \pm 2.18$ %

APPENDIX II

CALCULATION OF THEORETICAL CROSS-TALK RATIO

An approximation of the cross-talk ratio between adjacent channels may be made by assuming conservatively an RC type exponential decay. Since cross-talk is a function of the band-width, an equivalent time constant may be derived from the band-width of the system, which may be taken as the cut-off frequency of the output filter.

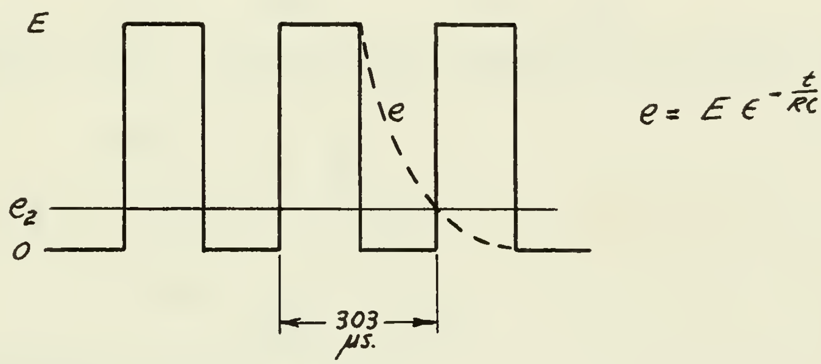

FIG. 20

$$\text{Cross-talk ratio} = 20 \log \frac{E}{e_2}$$

Figure 20 illustrates the assumed RC type exponential decay of a channel pulse.

The following calculations and illustrations in Figure 21 show that cross-talk is a function of the band-width.

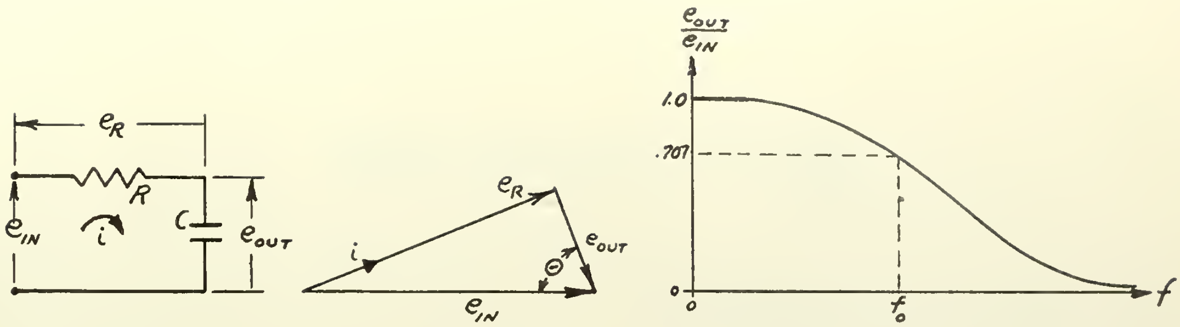

FIG. 21

$$\Theta = \text{phase angle between } e_{out} \text{ and } e_{in}$$

$$f_o \equiv \frac{1}{2\pi RC}$$

$$\tan\Theta = \frac{e_R}{e_{OUT}} = \frac{R}{X_c} = R\omega C = 2\pi RCf = \frac{f}{f_o}$$

$$\cos\Theta = \frac{e_{OUT}}{e_{IN}} = \cos\tan^{-1}\frac{f}{f_o}$$

At the 3 db. point on the response curve, $\Theta = 45°$

$$\tan\Theta = 1 = \frac{f}{f_o}$$

$$f = f_o = \frac{1}{2\pi RC}$$

$$RC = \frac{1}{2\pi f_o} = \frac{1}{\omega_o}$$

Referring to Figure 20:

$$e = E\epsilon^{-\frac{t}{RC}} = E\epsilon^{-2\pi f_o t} = E\epsilon^{-\omega_o t}$$

The cut-off frequency of the output filter = 3.6 kc.

$$f_o = 3.6 \text{ kc.} \quad , \quad \omega_o = 2.26 \times 10^4$$

$$t_2 = 1.515 \times 10^{-4}$$

$$e_2 = E\epsilon^{-3.43} = .032E$$

$$\text{Cross-talk ratio} = 20 \log \frac{E}{e_2}$$

$$= 20 \log 31.2 \cong 30 \text{ db.}$$